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(71) Applicant (*for all designated States except US*): **TELEFONAKTIEBOLAGET LM ERICSSON** [SE/SE]; S-126 25 Stockholm (SE).

(72) Inventor; and
(75) Inventor/Applicant (*for US only*): **HELLBERG, Richard** [SE/SE]; Forellvägen 14, 3tr, S-141 47 Huddinge (SE).

(74) Agents: **HEDBERG, Åke** et al.; Aros Patent AB, P.O. Box 1544, S-751 45 Stockholm (SE).

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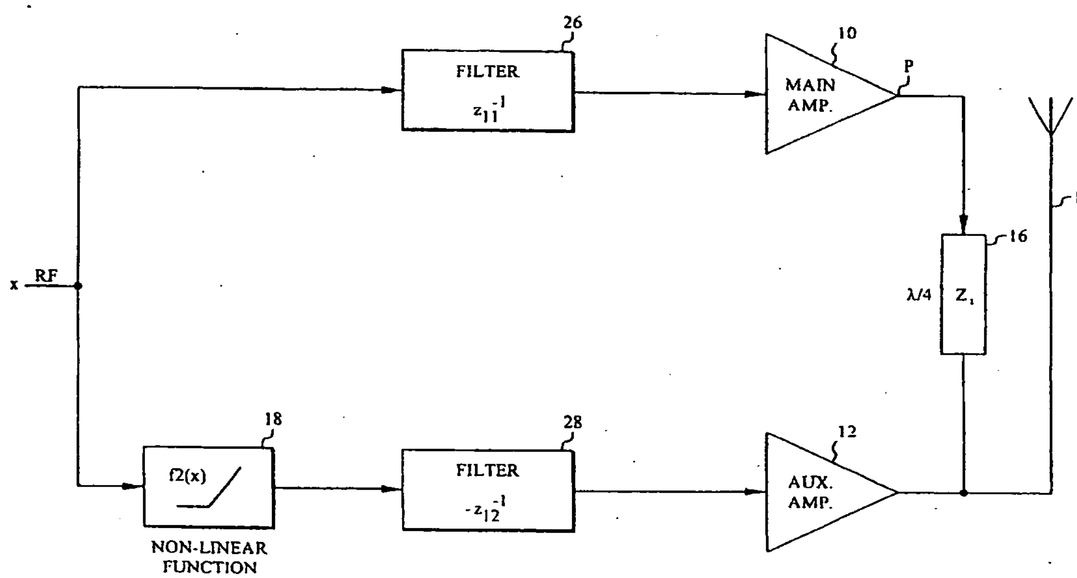
(54) Title: TRANSMITTER INCLUDING A COMPOSIT AMPLIFIER



(57) Abstract: A composite amplifier includes a main power amplifier (10) and an auxiliary power amplifier (12), which are connected to a load (14) over a Doherty output network (16). Filters (26, 28) are provided for pre-filtering the amplifier input signals in such a way that the signals meeting at the output of the main amplifier have essentially the same frequency dependence.

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For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

Transmitter including a composit amplifier

TECHNICAL FIELD

The present invention relates to a composite amplifier of the type that includes a main power amplifier and an auxiliary power amplifier, which are connected to a load over a Doherty output network. The invention also relates to a transmitter including such an amplifier and methods for operating such an amplifier and transmitter, respectively.

BACKGROUND

In cellular base stations, satellite communications and other communications and broadcast systems, many radio frequency (RF) carriers, spread over a large bandwidth, are amplified simultaneously in the same power amplifier. For the power amplifier this has the effect that the instantaneous transmit power will vary very widely and very rapidly. This is because the sum of many independent RF carriers (i.e. a multi-carrier signal) tends to have a large peak-to-average power ratio. It also tends to have a similar amplitude distribution as bandpass filtered Gaussian noise, which has a Rayleigh distribution.

A main difficulty in a power amplifier is efficiency. A conventional class B power amplifier exhibits maximum DC to RF power conversion efficiency when it delivers its peak power to the load. Since the quasi-Rayleigh distribution of amplitudes in the summed transmit signal implies a large difference between the average power and the peak power, the overall efficiency when amplifying such a signal in a conventional class B amplifier is very low. For a quasi-Rayleigh distributed signal with a 10 dB peak-to-average power ratio, the efficiency of an ideal class B amplifier is only 28%, see [1].

One way of increasing the efficiency of an RF power amplifier is to use the Doherty principle [1, 2, 3]. The Doherty amplifier uses in its basic form two amplifier stages, a main and an auxiliary amplifier (also called carrier and peaking amplifier, respectively). The load is connected to the auxiliary amplifier, and the main amplifier is connected to the load through an impedance-inverter, usually a quarter wavelength transmission line or an equivalent lumped network.

At low output levels only the main amplifier is active, and the auxiliary amplifier is shut off. In this region, the main amplifier sees a higher (transformed) load impedance than the impedance at peak power, which increases its efficiency in this region. When the output level climbs over the so-called transition point (usually at half the maximum output voltage), the auxiliary amplifier becomes active, driving current into the load. Through the impedance-inverting action of the quarter wavelength transmission line, this decreases the effective impedance at the output of the main amplifier, such that the main amplifier is kept at a constant (peak) voltage above the transition point. The result is a substantially linear output to input power relationship, with a significantly higher efficiency than a traditional amplifier.

The transition point can be shifted, so that the auxiliary amplifier kicks in at a lower or higher power level. This can be used for increasing efficiency for a specific type of signal or a specific amplitude distribution. When the transition point is shifted, the power division between the amplifiers at peak power is shifted accordingly, and the average power loss in each amplifier also changes. The latter effect also depends on the specific amplitude distribution.

An important feature of Doherty amplifiers is that they are inherently band-limited, since the impedance inverting network only provides 90 degrees of phase shift at a single frequency. This has the effect that the Doherty principle, i.e. the suppression of RF voltage rise at the main amplifier above

a certain transition point, works poorly (inefficiently) outside a limited frequency band. This is because the suppression requires the voltages from the main amplifier and the auxiliary amplifier to be in perfect anti-phase at the output of the main amplifier. Since the quarter-wave network is really only a quarter wave (90 degrees) phase shift at the center frequency, and shorter or longer at frequencies below and above the center frequency, respectively, this requirement gets more and more violated the further one gets from the center frequency of the impedance inverter.

SUMMARY

An object of the present invention is to enhance efficiency of a composite amplifier provided with a Doherty output network. Preferably the efficiency is increased over a broader frequency band.

The stated object is achieved in accordance with the attached claims.

Briefly, the present invention enhances efficiency by separately pre-filtering the input signals to the power amplifiers in such a way that the signals meeting at the output of the main amplifier have the same frequency dependence. Preferably this is done by using filters representing the inverses of the frequency dependent power amplifier impedance and transimpedance, thereby flattening the frequency response of the composite amplifier over a broader frequency band.

BRIEF DESCRIPTION OF THE DRAWINGS

The invention, together with further objects and advantages thereof, may best be understood by making reference to the following description taken together with the accompanying drawings, in which:

Fig. 1 is a simplified block diagram of a Doherty amplifier;

Fig. 2 is a model of the output network of a Doherty amplifier;

Fig. 3 is a diagram illustrating the frequency dependence of the trans-impedance between the auxiliary and main amplifier;

Fig. 4 is a simplified block diagram of an exemplary embodiment of the composite amplifier in accordance with the present invention;

Fig. 5 is a diagram illustrating the input-output voltage characteristics of a prior art Doherty amplifier;

Fig. 6 is a diagram illustrating the input-output voltage characteristics of a composite amplifier in accordance with the embodiment of fig. 4;

Fig. 7 is a simplified block diagram of another exemplary embodiment of the composite amplifier in accordance with the present invention;

Fig. 8 is a simplified block diagram of still another exemplary embodiment of the composite amplifier in accordance with the present invention;

Fig. 9 is a diagram illustrating the input-output voltage characteristics of a composite amplifier in accordance with the embodiment of fig. 7;

Fig. 10 is a diagram illustrating the input-output voltage characteristics of a composite amplifier in accordance with the embodiment of fig. 8;

Fig. 11 is a simplified block diagram of a further exemplary embodiment of the composite amplifier in accordance with the present invention;

Fig. 12 is a diagram illustrating the input-output voltage characteristics of a composite amplifier in accordance with the embodiment of fig. 11; and

Fig. 13 is a block diagram of an exemplary implementation of the embodiment of fig. 11.

DETAILED DESCRIPTION

Fig. 1 is a simplified block diagram of a Doherty amplifier. It includes a main power amplifier 10 and an auxiliary power amplifier 12. The output of auxiliary amplifier 12 is connected directly to a load (antenna) 14, whereas the output of main amplifier 10 is connected to the output of auxiliary amplifier 12 over a Doherty output network including a quarter wavelength transmission line 16. On the input side an RF (Radio Frequency) input signal x is divided into two branches, one branch intended for main amplifier 10 and another branch for auxiliary amplifier 12. The auxiliary amplifier branch

includes a non-linear function block 18, which transforms input signal x into $f2(x)$, and a phase shifter 20, which shifts the input signal to auxiliary amplifier 12 by 90 degrees. As indicated by antenna 14 the composite amplifier may be part of a transmitter, for example a transmitter in a base station in a cellular mobile radio communication system.

Fig. 2 is a model of the output network of a Doherty amplifier. In this model the active part of the amplifier transistor outputs are modeled as linear controlled current generators. The finite output conductances of the transistors, together with possible reactances, are lumped together as z_{p1} and z_{p2}, respectively. The impedances presented to each current generator output node are defined as:

$$z_{11} = \left.\frac{v_1}{i_1}\right|_{i_2 = 0} \qquad z_{22} = \left.\frac{v_2}{i_2}\right|_{i_1 = 0}$$

Similarly, the transimpedances, i.e. the voltage at the inactive amplifier output in response to an output current at the active amplifier, are defined as:

$$z_{21} = \left.\frac{v_2}{i_1}\right|_{i_2 = 0} \qquad z_{12} = \left.\frac{v_1}{i_2}\right|_{i_1 = 0}$$

Assuming that all components are reasonably linear, superposition can be used for analyzing this model. The composite amplifier output voltage (at the antenna) is here assumed to be the same as the output voltage at auxiliary amplifier 12, although in reality there can be a feeder cable, filters, etc. separating the actual antenna and the amplifier output. The combined effect of all these elements is included in the antenna (output) impedance, z_{ANT}.

In an ideal lossless Doherty amplifier the impedance z_{11} and the transimpedance z_{12} are both affected by a frequency-dependent reactive part due to the quarter-wave transformer, which is only a perfect quarter wavelength at a single frequency, as well as reactive components of z_{p1}, z_{p2} and z_{ANT}. However, the impedance z_{11} and the transimpedance z_{12} are also affected by losses due to the fact that the magnitude of the voltage at the opposite terminal is lowered for a given current stimulus. The frequency dependence of transimpedance z_{12} is illustrated in fig. 3 for both the lossless and lossy case (in fig. 3 the design frequency is 1 GHz). The impedance z_{11} would have a qualitatively similar frequency dependence.

Since the primary function of auxiliary amplifier 12 in a Doherty amplifier is to keep the voltage at main amplifier 10 below saturation, the frequency dependence of all signals at the output *P* of main amplifier 10 should be the same. Thus, the output signal from main amplifier 10 and the transformed (by Z_1) output signal from auxiliary amplifier 12, which meets the output signal from main amplifier 10 at *P* and keeps amplifier 10 below saturation, should have the same frequency dependence, and this frequency dependence should preferably be as flat as possible. The output *P* is located right at the power amplifier transistor collector. This can be achieved by pre-filtering the input signals to amplifiers 10 and 12 in such a way that the combined filtering actions of impedance and transimpedance is the same for both signals at output *P*.

Fig. 4 illustrates an exemplary embodiment of the present invention achieves this result. In this embodiment the frequency dependence of the output signal from main amplifier 10 is eliminated by filtering the input signal with a filter having the frequency characteristics of z_{11}^{-1}, the inverse filter of the impedance seen at the output of main amplifier 10. Similarly, an equalization of the transformed output signal from auxiliary amplifier 12 may obtained by filtering its input with a filter having the frequency characteristics of z_{12}^{-1}, the inverse filter of the transimpedance between auxiliary

amplifier 12 and main amplifier 10. The terms z_{11}^{-1} and z_{12}^{-1} are, when observed in the frequency domain, equal to $1/z_{11}$ and $1/z_{12}$, respectively.

The analytical expressions for obtaining i_1 and i_2 may be expressed as:

$$i_1 = \frac{V_{\max}}{\alpha} z_{11}^{-1} * x$$

$$i_2 = -V_{\max} z_{12}^{-1} * f2(x)$$

where f2(x) is a function that is 0 up to the transition point α, and thereafter has the same slope as x, as illustrated in block 18..

If the dimensionless signals f2(x) and x are represented in the time domain, "*" represent convolution in the time domain. If they are represented in the frequency domain, the symbol instead represents multiplication of frequency responses, and the multiplication with inverse filters can be written as a division by the filter instead. The derived network model is shown in fig. 4. Filters 26 and 28 may thus be represented by:

$$\text{Filter 26:} \quad \frac{V_{\max}}{\alpha} z_{11}^{-1}$$

$$\text{Filter 28:} \quad -V_{\max} z_{12}^{-1}$$

So far only the optimization of the voltage at the output *P* of main amplifier 10 has been studied, and expressions for the optimal currents have been derived. The voltage amplitude at auxiliary amplifier 12 has been left out of the discussion. This is partly because a fixed hardware setup has been assumed, i.e. the impedance of the quarter wave line and the load has been assumed fixed. For a lossless system this is not a serious problem, the effect of optimizing for flat response and optimal amplitude at main amplifier 10 is that the output signal gets a slight frequency dependence. When losses are considered, however, the effect can be that the maximum voltage at auxiliary

amplifier 12 never reaches V_{max}, even at maximum input levels. This constitutes a more serious problem, since the transistors then deliver less than the maximum power to the load (at peak output), while still having the same supply voltage, and the efficiency will drop. The simple solution is to either reduce the supply voltage, or to increase the load impedance until maximum voltage is achieved at peak output (the latter solution is preferred, since this scheme gives higher efficiency and more available output power). The compensation for losses can also have the effect that neither transistor reaches I_{max}, which also implies an under-utilization of the transistors. Impedances (load and quarter-wave line) may then have to be changed in order to use the maximum possible output power from the transistors. Equally important is to keep both transistors in the safe region, so that the maximum currents and voltages are reached but not exceeded. Note that when changing the impedances in the circuit, redesign of the compensations according to the depicted scheme is necessary. Also, if maximum power is not a design goal, the circuit can be optimized differently, to meet other objectives.

The effect of the compensation in accordance with the embodiment of fig. 4 is illustrated in fig. 5 and 6 with reference to a simulated example with a multi-carrier signal.

In fig. 5 the normalized magnitudes of the voltages at main amplifier 10 and auxiliary amplifier 12 are plotted against the desired magnitude (the normalized amplitude of x) for the uncompensated case (prior art). The drive signals have been adjusted to keep both voltages within the linear (unsaturated) range of the transistors. The different slopes of the output signal (voltage at auxiliary amplifier 12) below and above the transition point indicate a static non-linearity. The different widths of these curves indicate a level-variant frequency dependence. The voltage at main amplifier 10 is not at all close to the desired constant level above the transition point, which means that the average efficiency will be low (although still probably better than for a class B amplifier).

The normalized magnitudes of the voltages at main amplifier 10 and auxiliary amplifier 12 after efficiency-boosting in accordance with the embodiment of fig. 4 are illustrated in fig. 6. Compensation of the network for losses has been performed by changing the transmission line impedance and the load impedance.

In the embodiment described with reference to fig. 4, the frequency dependence of the two signals meeting at output P was eliminated by equalizing filters on the input side. However, equalization is not strictly necessary from an efficiency-boosting viewpoint. Instead the essential feature is that both signals have the same frequency dependence. There are other ways to accomplish this. Two examples are illustrated in fig. 7 and 8. In both examples only one input signal is subjected to actual filtering, while the other input signal is only rescaled by a constant gain.

In the embodiment of fig. 7 the auxiliary amplifier branch is equalized by z_{12}^{-1} and then filtered by z_{11} to obtain the same frequency dependence as the main amplifier branch, which is only adjusted by a constant gain. Thus filters 26 and 28 are:

$$\text{Filter 26:} \quad \frac{V_{\max} \cdot k}{\alpha}$$

$$\text{Filter 28:} \quad -V_{\max} \cdot k \cdot z_{11} * z_{12}^{-1}$$

where k is a constant that is selected to make the voltage at the output node of main amplifier 10 equal to V_{max}.

In the embodiment of fig. 8 the main amplifier branch is equalized by z_{11}^{-1} and then filtered by z_{12} to obtain the same frequency dependence as the auxiliary amplifier branch, which is only adjusted by a constant gain. Thus filters 26 and 28 are:

$$\text{Filter 26:} \quad \frac{V_{max} \cdot k}{\alpha} z_{12} * z_{11}^{-1}$$

$$\text{Filter 28:} \quad -V_{max} \cdot k$$

Fig. 9 and 10 illustrate the input-output voltage characteristics of the embodiments of fig. 7 and 8, respectively. As can be seen from these figures, the voltage at main amplifier 10 is close to the desired constant level above the transition point, which means that the average efficiency will be high. As expected, the lines are somewhat widened compared to fig. 6 due to the reduced equalization.

In the embodiments described with reference to fig. 7 and 8, the frequency dependence of the two signals meeting at output *P* was partly reduced by equalizing filters on the input side (z_{12}^{-1} and z_{11}^{-1}, respectively). However, as noted above, equalization is not strictly necessary from an efficiency-boosting viewpoint. Thus, by considering only the essential feature that both signals should have the same frequency dependence, it is possible to eliminate inverse filters. An examples is illustrated in fig. 11. In this example both input branches include filters that emulate the filtering produced by the other branch, thereby subjecting each branch to the same total filter.

Thus, in the embodiment of fig. 11 the main amplifier branch is filtered by a filter having the frequency characteristics of z_{12}, while the auxiliary amplifier branch is filtered by a filter having the frequency characteristics of z_{11}. Thus filters 26 and 28 are:

$$\text{Filter 26:} \quad \frac{V_{max} \cdot k}{\alpha} z_{12}$$

$$\text{Filter 28:} \quad -V_{max} \cdot k \cdot z_{11}$$

Fig. 12 illustrates the input-output voltage characteristics of the embodiment of fig. 11. As can be seen from this figure, the voltage at main amplifier 10 is close to the desired constant level above the transition point, which means that the average efficiency will be high also in this embodiment. As expected, the lines are somewhat more widened compared to fig. 9 and 10 due to the eliminated equalization.

Fig. 13 is a block diagram of an exemplary implementation of the embodiment of fig. 11. A simple but elegant method for obtaining the filters z_{11} and z_{12} (the filtering by z_{12} can be obtained by using z_{21} instead) is to use input-side copies of the Doherty output network, containing the same passive circuit elements that are present in the actual output network. When such a network is driven by a current generator (small-signal transistor) on the input side, the output voltage automatically has the right frequency dependence. The requirement for this to work is that the transistor output parasitic elements, the quarter-wave line and the antenna network impedance can be accurately modeled. A possibility is to scale the impedance of all elements in the network to get more realizable values and/or better voltage and current levels.

If the non-linear function f2(x) of the RF signal is produced by a class C amplifier, it can also be produced by driving amplifier G3 in class C mode. The amplification to higher voltage is preferably done in the preamplifiers to main amplifier 10 and auxiliary amplifier 12. The antenna network impedance is in this case modeled by a 50 Ohm resistance with a parallel resonator tuned to the center frequency. Amplifier G3 is a controlled current generator. The input impedance of (identical) amplifiers G2 and G4 together with appropriate additional reactances emulate the antenna network impedance Z_{ANT}, and possible parasitics on the output of G3 are included in the corresponding Z_{p2} and Z_{p1}. Amplifier G1 provides a matching gain to main amplifier 10. The power amplifiers include necessary input matching networks and preamplifiers.

In practice, the performance of the described methods will depend on how well the characteristics of the Doherty output network are known. Measuring transimpedances in the output network is often hard to do directly, since the (RF) voltage probe and the current injector will always have parasitics that must be taken into account. Indirectly, impedance parameters (Z-parameters) can be extracted by traveling wave measurements (S-parameters). A combination of different parameters that are easy to measure can also be selected. The required filters or emulating networks can then be designed using extracted impedances and transimpedances.

The gain of the linear path to main amplifier 10 can be adjusted (at several frequencies to ensure amplitude flatness) by observing the starting point of compression in the output for a main. Compression should occur at a power corresponding to the transition point, if f2(x) is deactivated.

Optimal suppression of the voltage rise at main amplifier 10 above the transition point, requires phase and gain matching of the linear part to the non-linear part at this node. The phase matching, or electrical path length difference, should be sufficiently correct (within a fraction of a wavelength) before adjustment in order to avoid local minima at multiple wavelengths away from the correct one.

Probing the voltage at main amplifier 10 for flatness above the transition point, instead of just observing the efficiency, can help in achieving maximum efficiency. The probe must have high impedance to avoid increasing the losses or otherwise detrimentally affect the conditions in the circuit. Apart from that, the probe impedance can be incorporated in the efficiency-boosting compensations.

Many parameters of the output network and amplifiers are slowly changing, due to aging, temperature variations and other environmental changes. This means that the efficiency of the amplifier may degrade from its initial level.

To cope with this problem, the filters and gains in the network can be made to respond in real time to the parameter variations.

The adjustments described in the previous paragraphs can be automated by monitoring the output and possibly the voltage at main amplifier 10 and relate this to the signals inputted at various points in the network. The measured values can then be used for changing the parameters of the input network. An alternative is to insert special signals that are only used for measurements (pilot signals).

An entirely digital implementation of the efficiency-boosting techniques will have advantages over an analog implementation in that the filters will be more accessible to detailed adjustments. An analog implementation relies on the adjustments of circuit elements, but the circuit itself is hard to change during operation.

Since saturation is a somewhat vaguely defined state, with a transition region in which the power amplifier is neither a pure current source nor a hard-limited voltage source, solutions can be found in which a power amplifier is held slightly saturated over the “flat” voltage range. The methods proposed by the present invention can be used to control this amount of saturation very precisely so that efficiency is increased, over that of a strictly non-saturated amplifier, but the distortion does not grow above a set limit.

Many different implementations are possible. Digital or analog signal processing can be used, and the processing can be performed with a variety of techniques, at baseband, intermediate or final (RF) frequencies. Arbitrary combinations of these can be used, matching the requirements for a function with a convenient way of implementing it. The solution can be used statically, optimized at the time of manufacture or at specific times during maintenance, or dynamically adaptive, for continuously optimizing the efficiency of the amplifier.

It will be understood by those skilled in the art that various modifications and changes may be made to the present invention without departure from the scope thereof, which is defined by the appended claims.

REFERENCES

[1] F. H. Raab, "Efficiency of Doherty RF Power Amplifier Systems", IEEE Trans. Broadcasting, vol. BC-33, no. 3, pp. 77-83, Sept. 1987.

[2] US Patent No. 5.420,541 (D. M. Upton et al.).

[3] US Patent No. 5,568,086 (J.J. Schuss et al.).

CLAIMS

1. A composite amplifier including

a main power amplifier;

an auxiliary power amplifier, said amplifiers being connected to a load over a Doherty output network; and

means for pre-filtering amplifier input signals in such a way that the signals meeting at the output of said main amplifier have essentially the same frequency dependence.

2. The composite amplifier of claim 1, including a pre-filter to said main amplifier having the same frequency dependence as the inverse of the impedance presented to the main amplifier current generator output node, and a pre-filter to said auxiliary amplifier having the same frequency dependence as the inverse of the transimpedance between said auxiliary amplifier and said main amplifier.

3. The composite amplifier of claim 1, including a pre-filter to said auxiliary amplifier having the same frequency dependence as a filter combination formed by the inverse of the impedance presented to the main amplifier current generator output node and the inverse of the transimpedance between said auxiliary amplifier and said main amplifier.

4. The composite amplifier of claim 1, including a pre-filter to said main amplifier having the same frequency dependence as a filter combination formed by the transimpedance between said auxiliary amplifier and said main amplifier and the inverse of the inverse of the impedance presented to themain amplifier current generator output node.

5. The composite amplifier of claim 1, including a pre-filter to said main amplifier having the same frequency dependence as the transimpedance between said auxiliary amplifier and said main amplifier, and a pre-filter to said auxiliary amplifier having the same frequency dependence as the inverse

of the impedance presented to the main amplifier current generator output node.

6. A transmitter with a composite amplifier including

a main power amplifier;

an auxiliary power amplifier, said amplifiers being connected to a load over a Doherty output network; and

means for pre-filtering amplifier input signals in such a way that the signals meeting at the output of said main amplifier have essentially the same frequency dependence.

7. The transmitter of claim 6, including a pre-filter to said main amplifier having the same frequency dependence as the inverse of the inverse of the impedance presented to the main amplifier current generator output node, and a pre-filter to said auxiliary amplifier having the same frequency dependence as the inverse of the transimpedance between said auxiliary amplifier and said main amplifier.

8. The transmitter of claim 6, including a pre-filter to said auxiliary amplifier having the same frequency dependence as a filter combination formed by the inverse of the impedance presented to the main amplifier current generator output node and the inverse of the transimpedance between said auxiliary amplifier and said main amplifier.

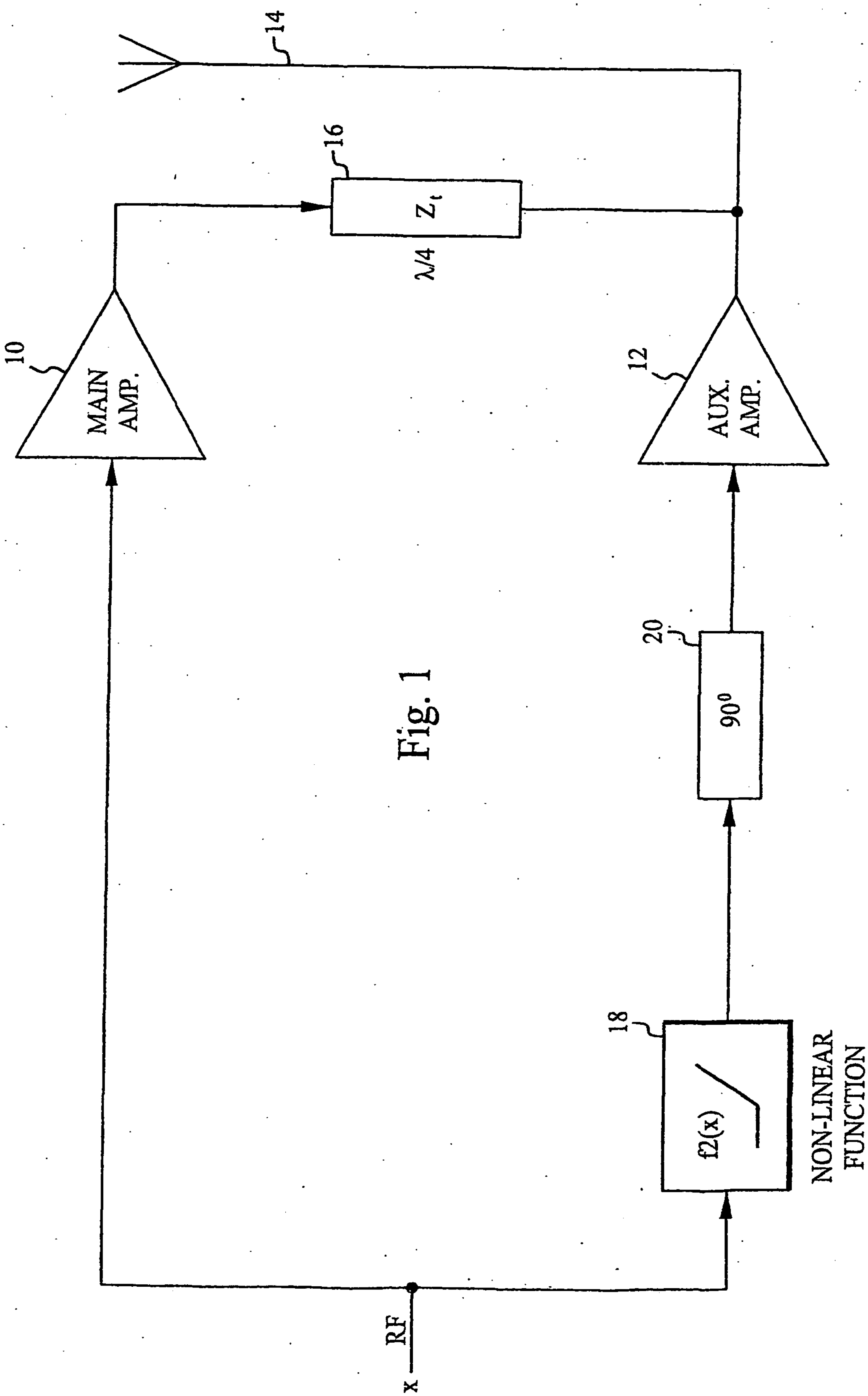
9. The transmitter of claim 6, including a pre-filter to said main amplifier having the same frequency dependence as a filter combination formed by the transimpedance between said auxiliary amplifier and said main amplifier and the inverse of the inverse of the impedance presented to the main amplifier current generator output node.

10. The transmitter of claim 6, including a pre-filter to said main amplifier having the same frequency dependence as the transimpedance between said

auxiliary amplifier and said main amplifier, and a pre-filter to said auxiliary amplifier having the same frequency dependence as the inverse of the impedance presented to the main amplifier current generator output node.

11. A method of operating a composite amplifier including a main power amplifier and an auxiliary power amplifier, which are connected to a load over a Doherty output network, including the step of pre-filtering amplifier input signals in such a way that the signals meeting at the output of said main amplifier have essentially the same frequency dependence.

12. A method of operating a transmitter provided with a composite amplifier including a main power amplifier and an auxiliary power amplifier, which are connected to a load over a Doherty output network, including the step of pre-filtering amplifier input signals in such a way that the signals meeting at the output of said main amplifier have essentially the same frequency dependence.


14
16
Z_t
$\lambda/4$
10
MAIN AMP.
12
AUX. AMP.
20
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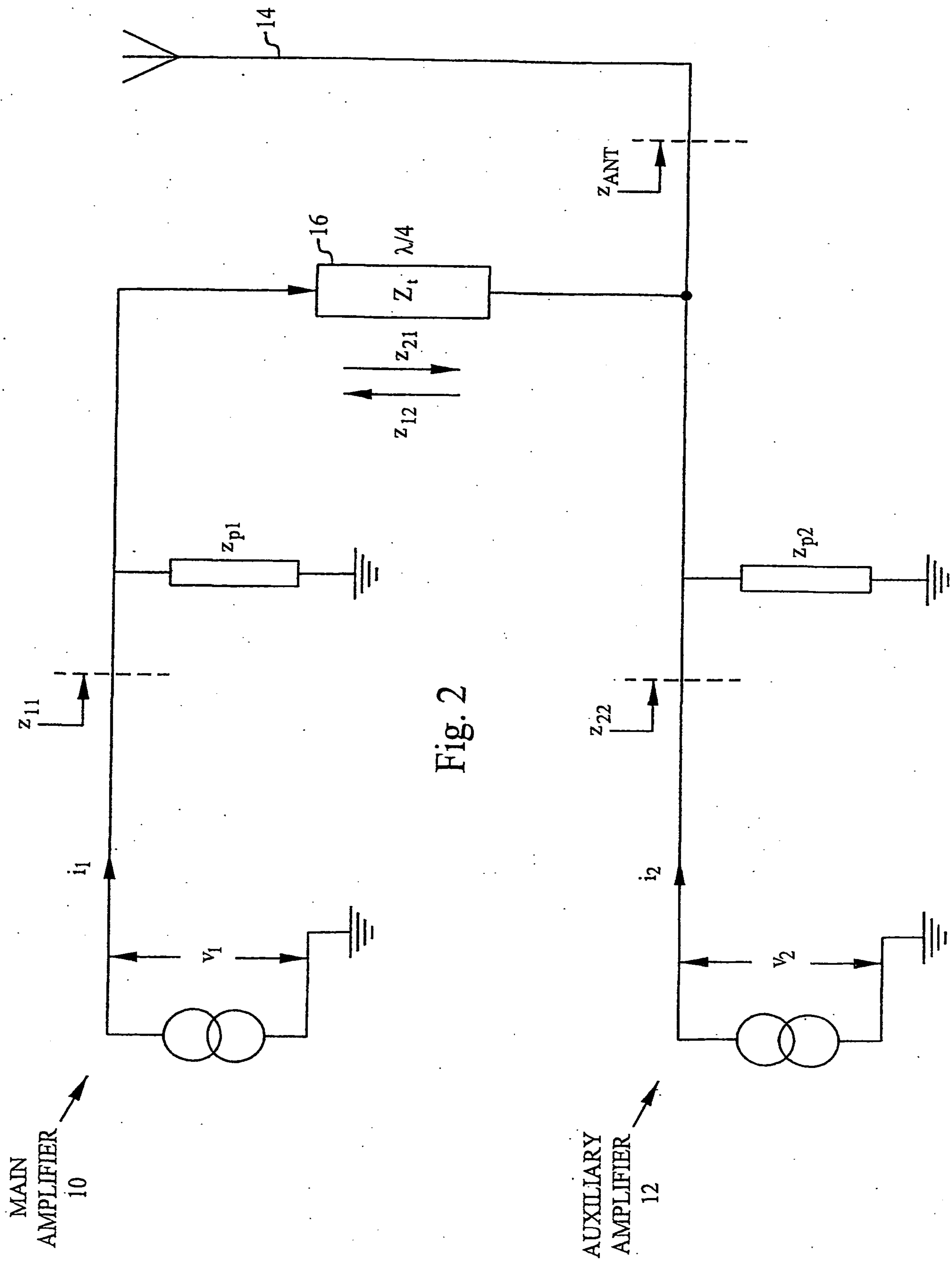
18
f2(x)
NON-LINEAR FUNCTION
RF
x

Fig. 1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λ/4
Zt
Z21
Z12
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Z11
Z22
i1
i2
v1
v2
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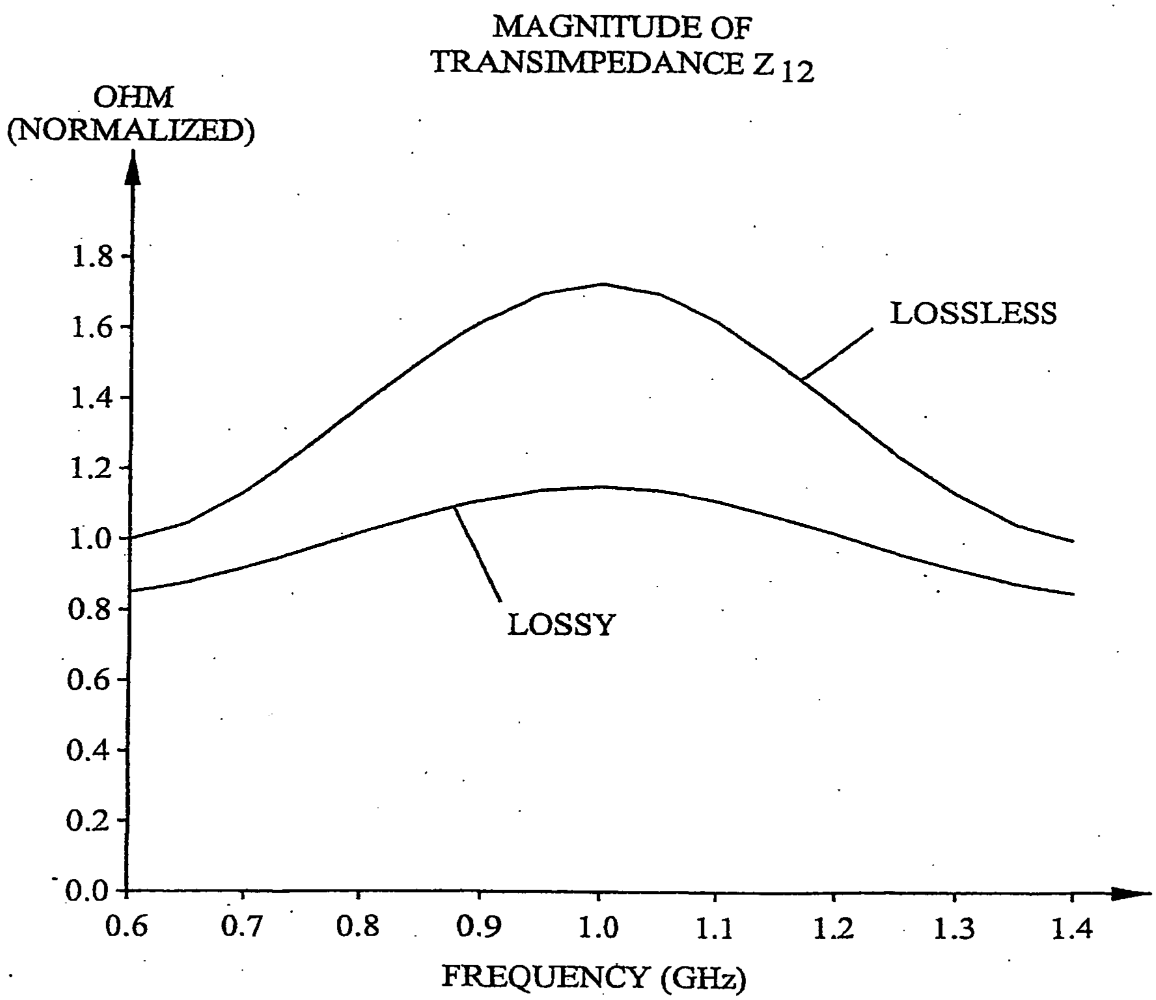
AUXILIARY AMPLIFIER 12

Fig. 2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0.6
0.4
0.2
0.0
LOSSLESS
LOSSY
0.6
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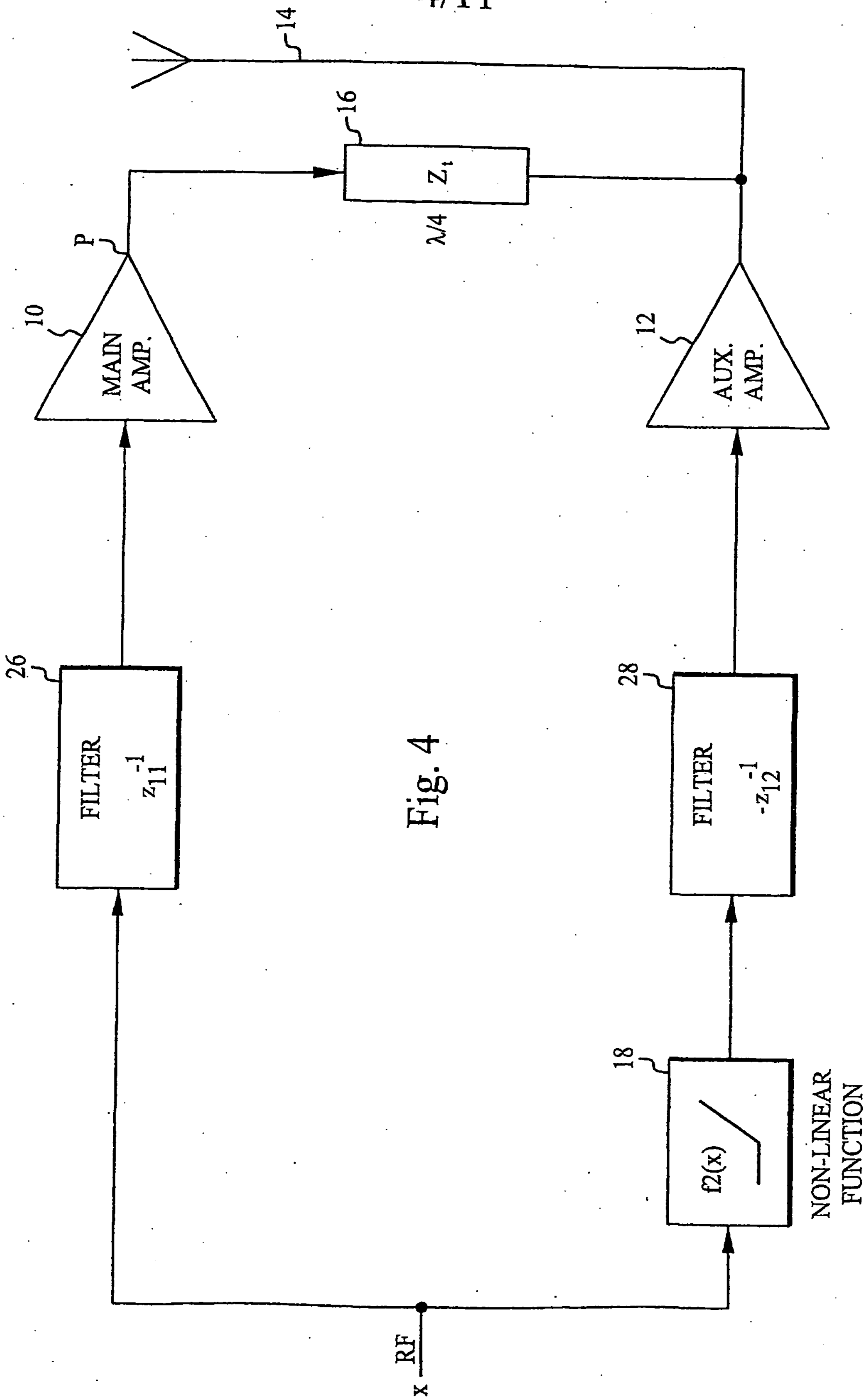
0.8
0.9
1.0
1.1
1.2
1.3
1.4
FREQUENCY (GHz)

Fig. 3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16
Z_t
λ/4
P
10
MAIN AMP.
12
AUX. AMP.
26
FILTER z_{11}^{-1}
28
FILTER $-z_{12}^{-1}$
18
f2(x)
NON-LINEAR FUNCTION
RF
x

Fig. 4

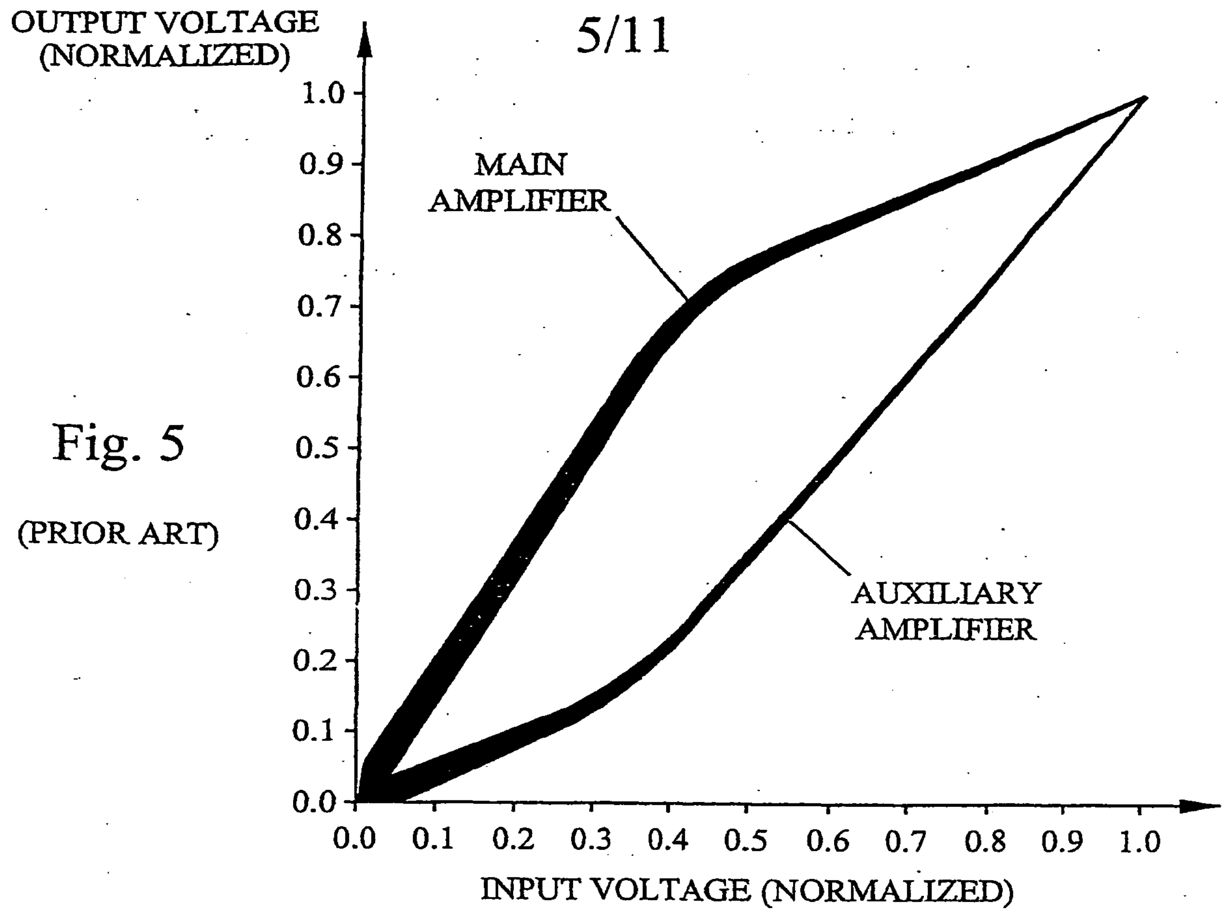


Fig. 5

(PRIOR ART)

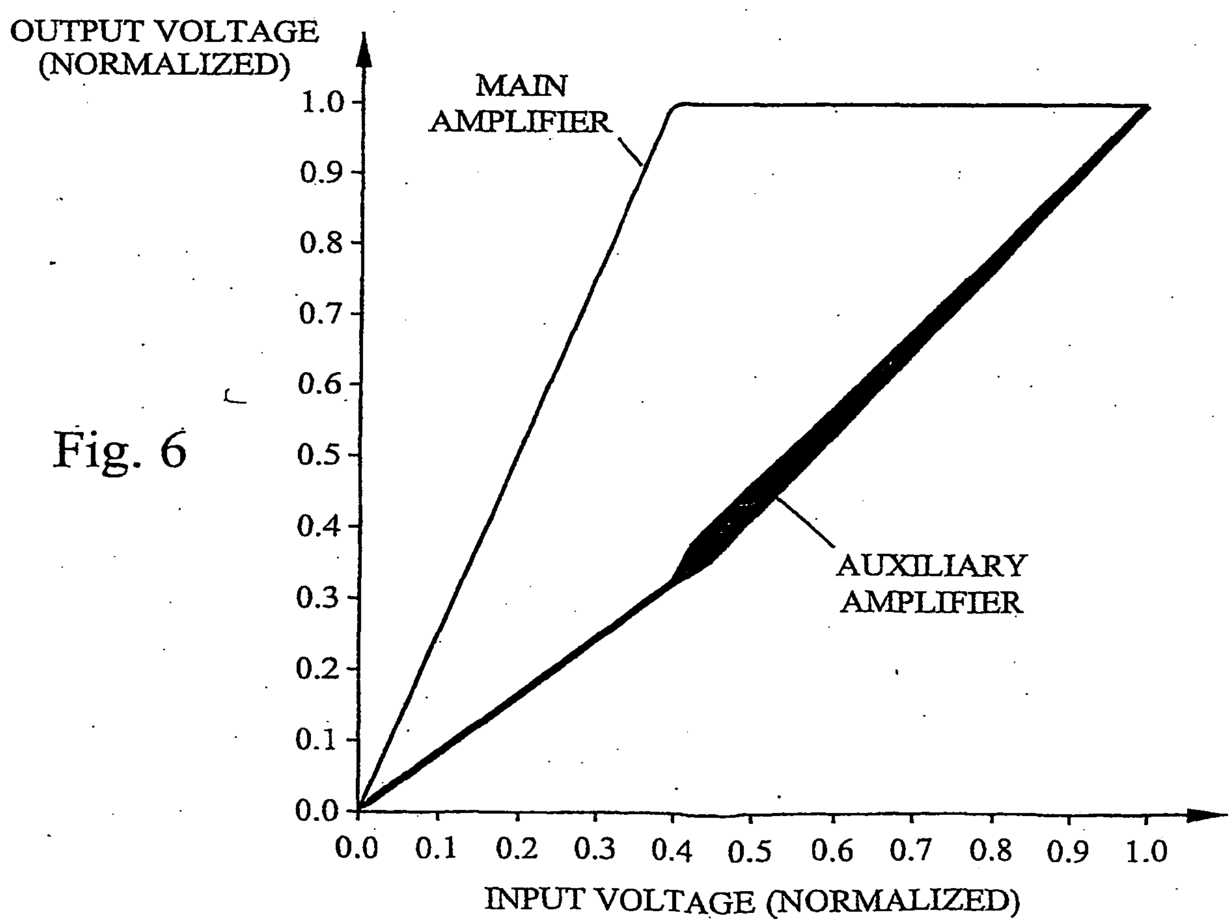


Fig. 6

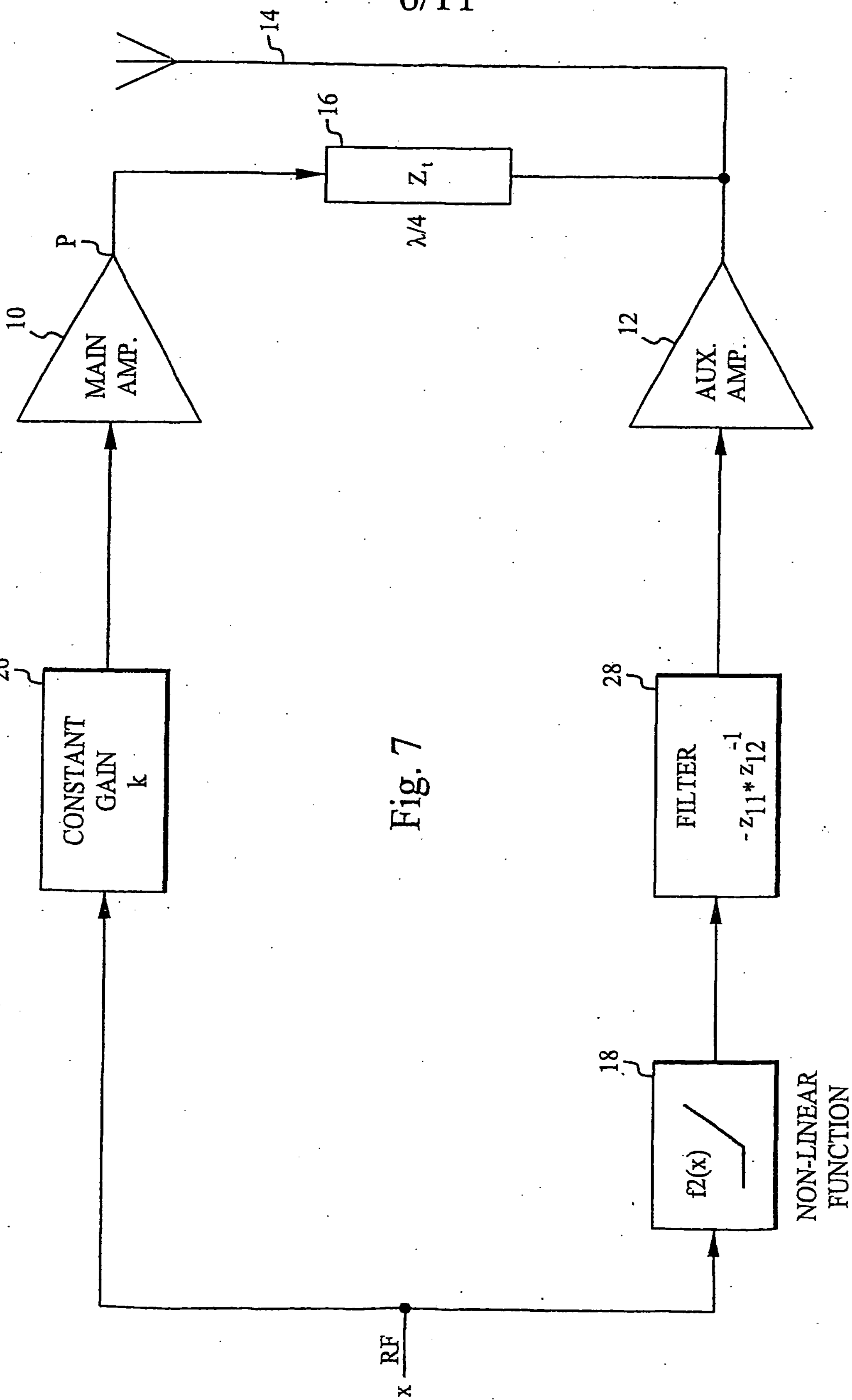

14
16
Z_t
$\lambda/4$
P
10
MAIN AMP.
12
AUX. AMP.
26
CONSTANT GAIN k
28
FILTER
$-z_{11} * z_{12}^{-1}$
18
f2(x)
NON-LINEAR FUNCTION
RF
x

Fig. 7

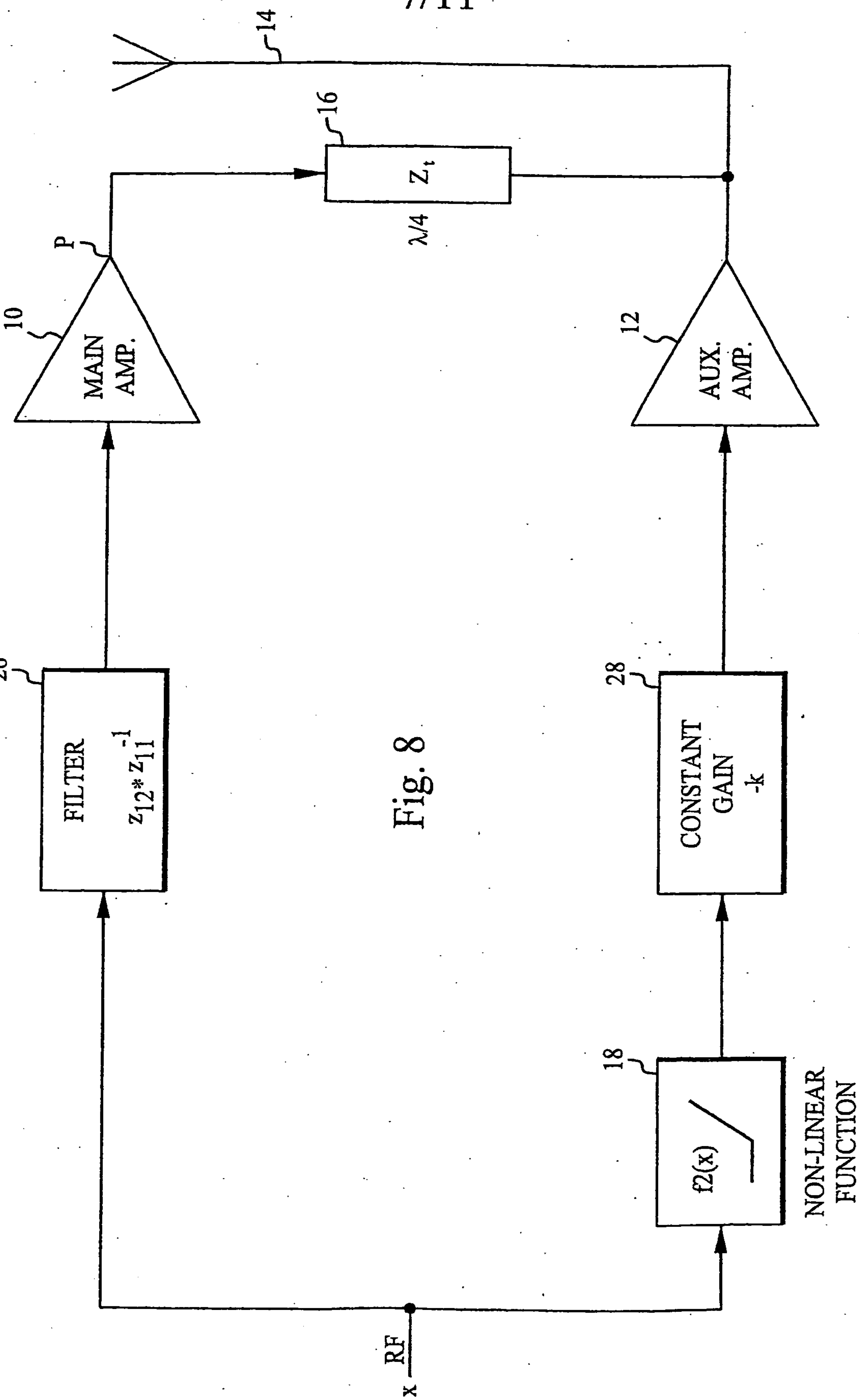

14
16
Z_t
$\lambda/4$
P
10
MAIN AMP.
12
AUX. AMP.
26
FILTER
$z_{12} * z_{11}^{-1}$
28
CONSTANT GAIN
-k
18
f2(x)
NON-LINEAR FUNCTION
RF
x

Fig. 8

Fig. 9

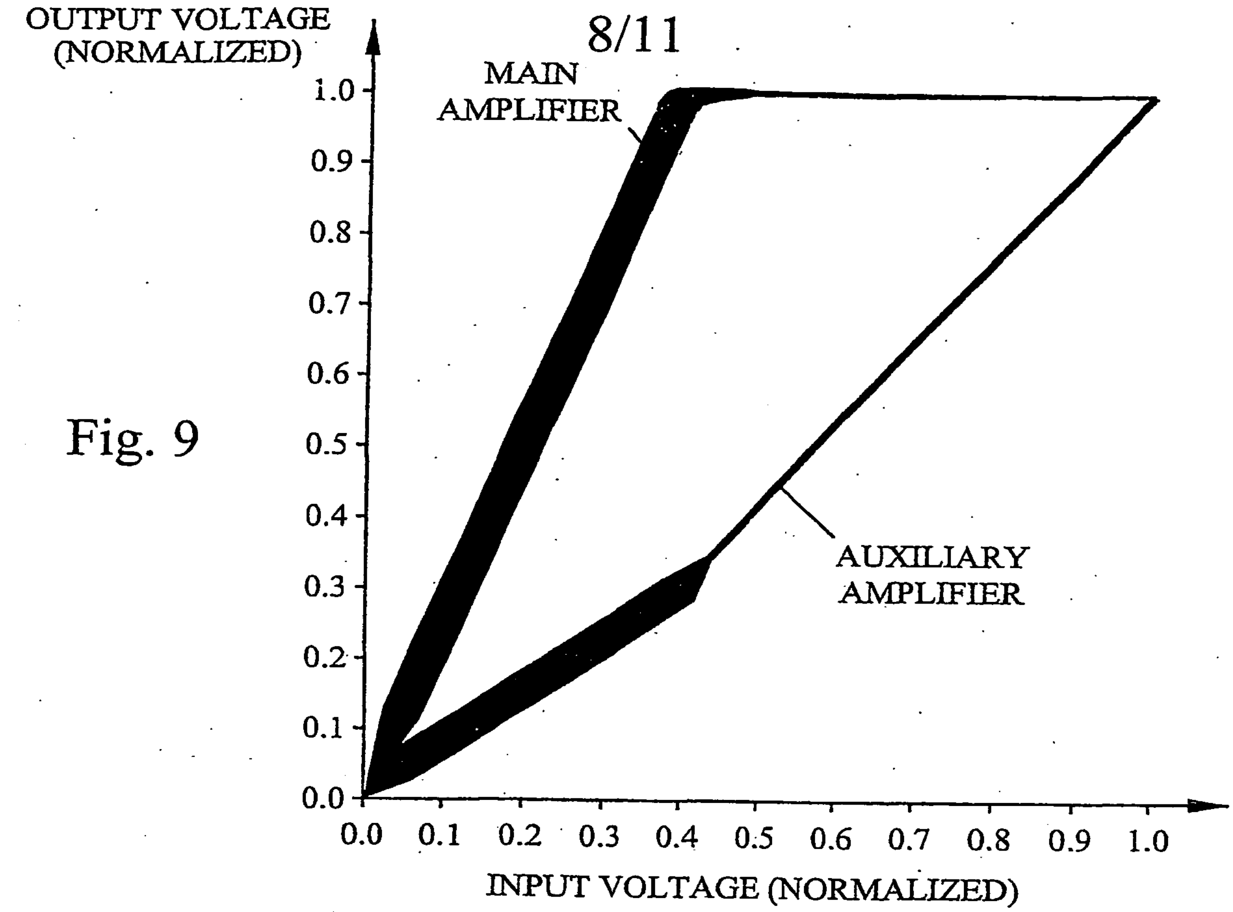
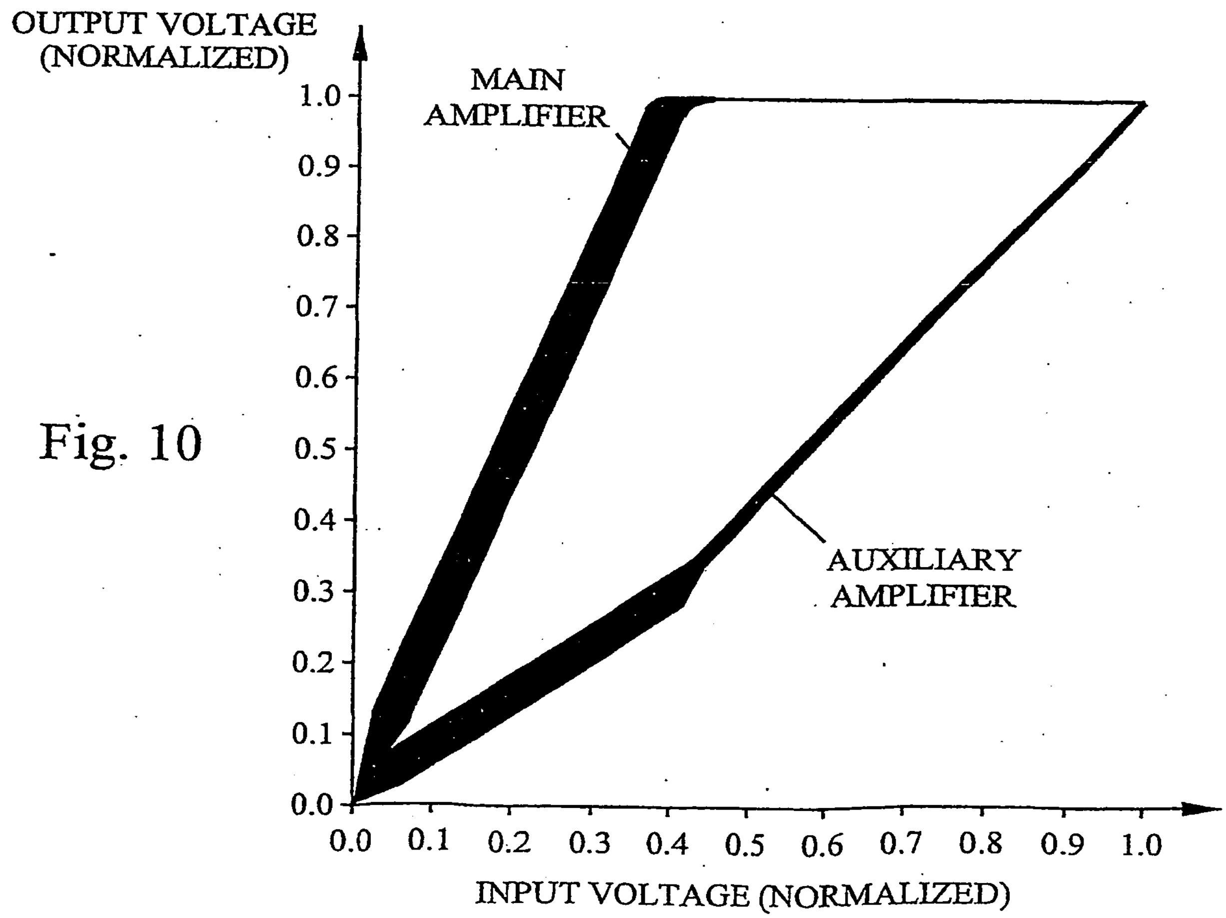

OUTPUT VOLTAGE
(NORMALIZED)
MAIN
AMPLIFIER
AUXILIARY
AMPLIFIER
1.0
0.9
0.8
0.7
0.6
0.5
0.4
0.3
0.2
0.1
0.0
0.0 0.1 0.2 0.3 0.4 0.5 0.6 0.7 0.8 0.9 1.0
INPUT VOLTAGE (NORMALIZED)

Fig. 10


OUTPUT VOLTAGE
(NORMALIZED)
MAIN
AMPLIFIER
AUXILIARY
AMPLIFIER
1.0
0.9
0.8
0.7
0.6
0.5
0.4
0.3
0.2
0.1
0.0
0.0 0.1 0.2 0.3 0.4 0.5 0.6 0.7 0.8 0.9 1.0
INPUT VOLTAGE (NORMALIZED)

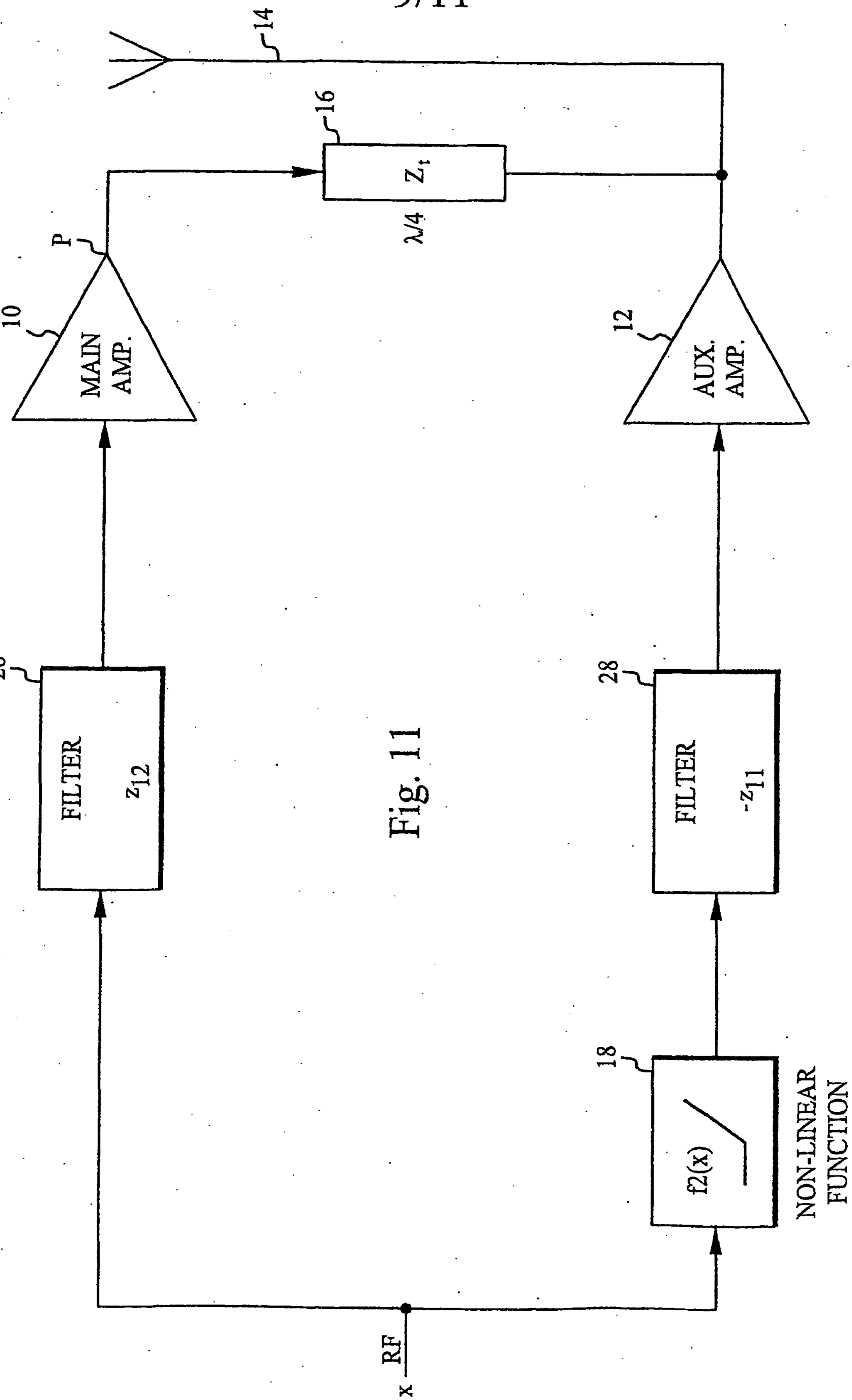

14
16
Z_t
λ/4
P
10
MAIN AMP.
12
AUX. AMP.
26
FILTER
z12
28
FILTER
-z11
18
f2(x)
NON-LINEAR FUNCTION
RF
x

Fig. 11

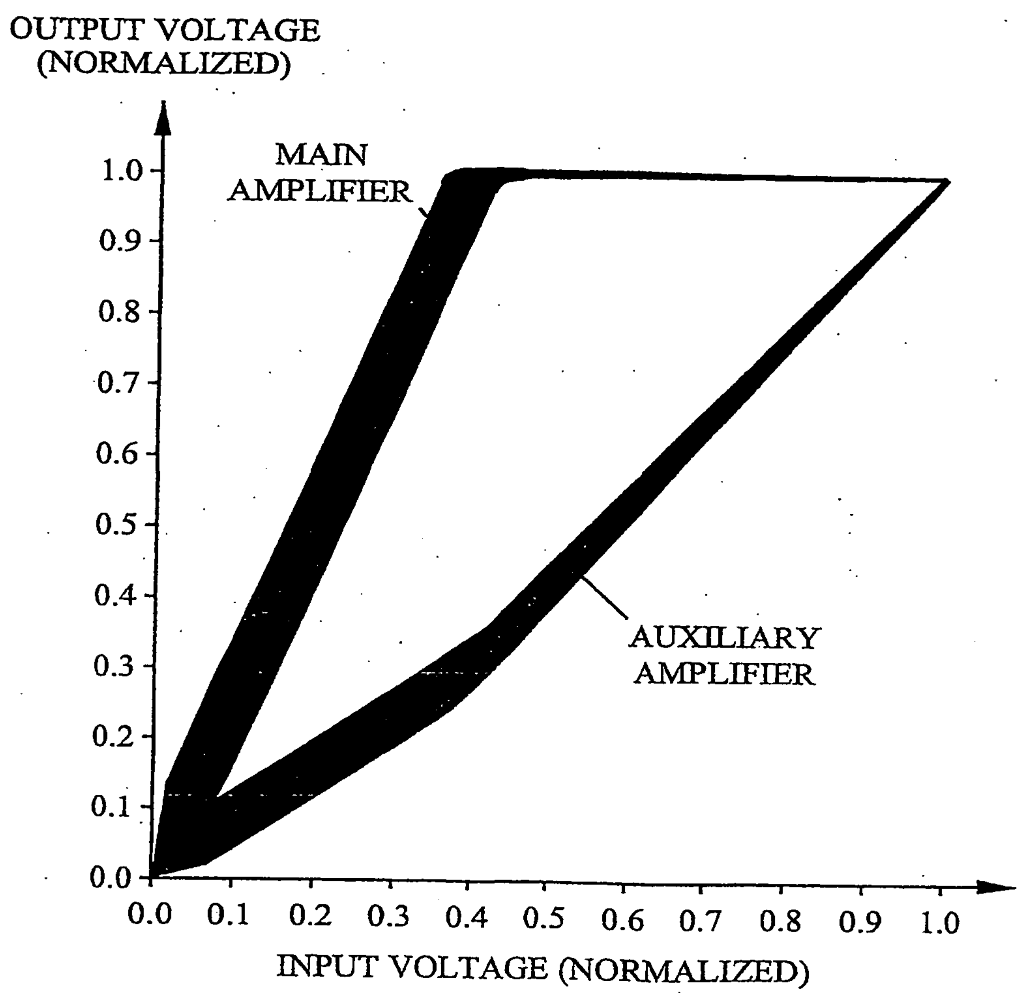

OUTPUT VOLTAGE
(NORMALIZED)
MAIN
AMPLIFIER
1.0
0.9
0.8
0.7
0.6
0.5
0.4
0.3
0.2
0.1
0.0
AUXILIARY
AMPLIFIER
0.0 0.1 0.2 0.3 0.4 0.5 0.6 0.7 0.8 0.9 1.0
INPUT VOLTAGE (NORMALIZED)

Fig. 12

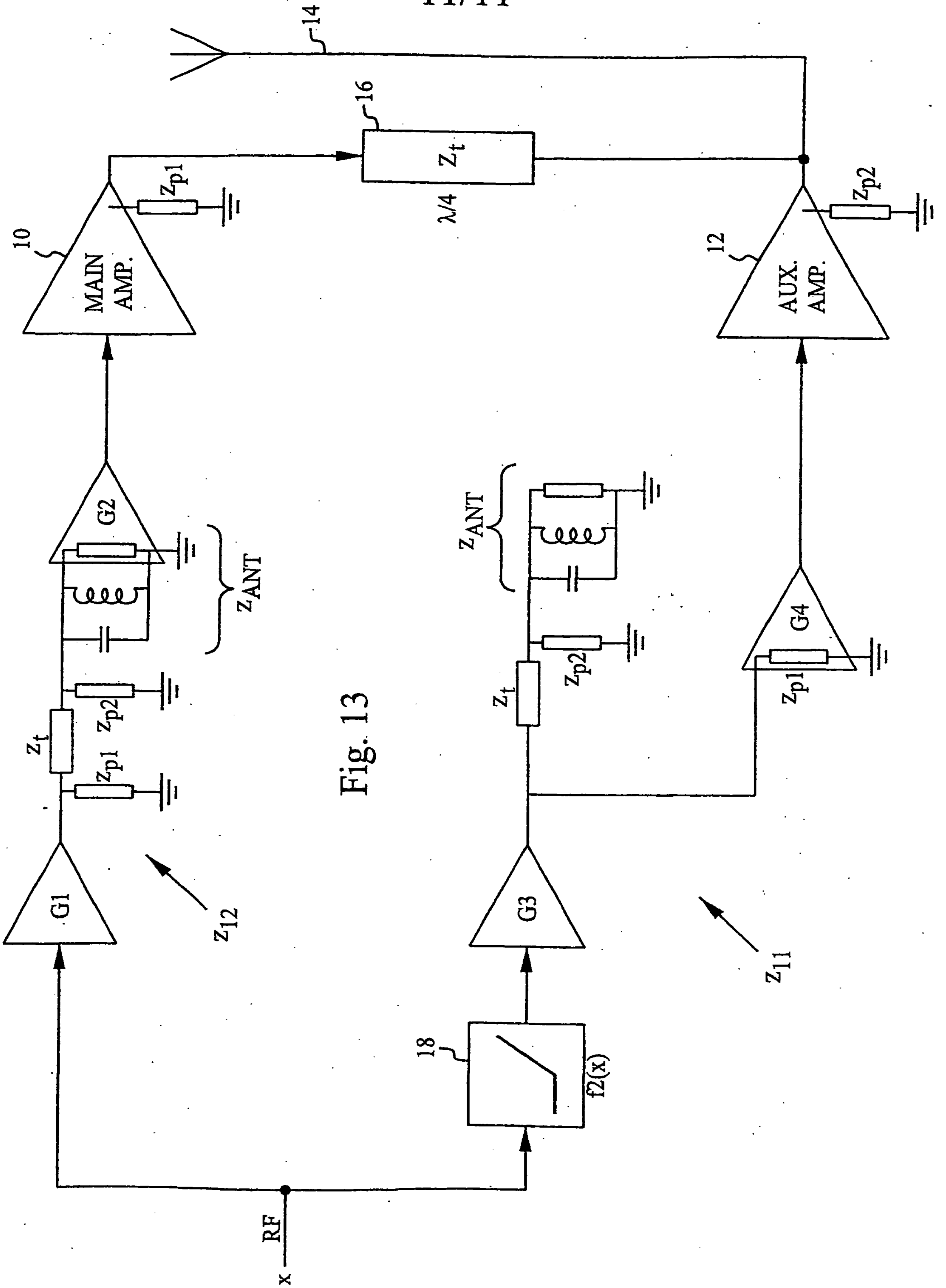

14
16
Zt
λ/4
Zp1
Zp2
10
MAIN AMP.
12
AUX. AMP.
G1
G2
G3
G4
ZANT
ZANT
Zt
Zp1
Zp2
Zt
Zp2
Zp1
Z12
Z11
18
f2(x)
RF
x

Fig. 13

INTERNATIONAL SEARCH REPORT

International application No.
PCT/SE 01/01392

A. CLASSIFICATION OF SUBJECT MATTER

IPC7: H03F 1/02, H04B 1/02

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

IPC7: H03F, H04B

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

SE,DK,FI,NO classes as above

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

WPI-DATA, EPO-INTERNAL, PAJ, INSPEC

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US 6097252 A (BERNARD EUGENE SIGMON ET AL), 1 August 2000 (01.08.00), column 3, line 55 - column 5, line 19, figure 1 --	1-5
A	US 6008694 A (EL-BADAWY AMIEN EL-SHARAWY), 28 December 1999 (28.12.99), column 2, line 38 - column 3, line 7, figure 1 --	1-5
A	US 6085074 A (LAWRENCE F. CYGAN), 4 July 2000 (04.07.00), column 3, line 49 - column 8, line 18 --	1-12
A	US 6128478 A (KEVIN KIM), 3 October 2000 (03.10.00), column 2, line 66 - column 7, line 7 --	1-12

☐ Further documents are listed in the continuation of Box C. ☒ See patent family annex.

* Special categories of cited documents:

"A" document defining the general state of the art which is not considered to be of particular relevance

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"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance: the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance: the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
8 October 2001	11 -10- 2001
Name and mailing address of the ISA/ Swedish Patent Office Box 5055, S-102 42 STOCKHOLM Facsimile No. +46 8 666 02 86	Authorized officer Antonio Farieta/MN Telephone No. +46 8 782 25 00

INTERNATIONAL SEARCH REPORT
Information on patent family members

03/09/01

International application No.
PCT/SE 01/01392

Patent document cited in search report	Publication date	Patent family member(s)	Publication date
US 6097252 A	01/08/00	WO 9856107 A	10/12/98
US 6008694 A	28/12/99	AU 4567999 A	01/02/00
		EP 1095448 A	02/05/01
		WO 0003480 A	20/01/00
		WO 0039138 A	06/07/00
US 6085074 A	04/07/00	NONE	
US 6128478 A	03/10/00	NONE	